Design of *K*-band slug tuners

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Abstract: An analysis of dual-slug line tuners is presented for two cases: distributed-transmission-line slugs and lumped-capacitance slugs. The effect of loss on tuner performance in the case of the transmission-line model is investigated and the curves obtained are used in the design of a *K*-band tuner over the frequency range 12–26 GHz. Good agreement is obtained between predicted and measured results.

1 Introduction

Although double-slug tuners have now been in use for over 30 years, [1] the only theoretical explanation published as to their working seems to be that of Paucksh *et al.* [2] In a similar manner, although metallic discs on microstrip lines are now commonly used to match GaAs MESFET transistors [3–5] disc dimensions have been selected on an entirely intuitive basis. An analysis of slug-transmission-line and microstrip disc tuners has recently been advanced by one of the authors, [6] but this study is based on simplifying assumptions which cannot always be justified.

In this communication, formulas are developed which allow the capabilities and the limitations of slug transmission-line tuners to be described in a remarkably simple manner. The effect of losses are also considered as these seriously limit the maximum values of voltage standing-wave ratios (VSWR) that may be obtained from a tuner, particularly at high frequencies.

The theory developed is applied to the design of a *K*-band tuner operating in the 12–26 GHz frequency range. The tuner is realised in a 4 mm diameter transmission line with slugs that are magnetically guided. Good agreement is obtained between the reflection coefficients measured on the prototype tuner and those theoretically predicted.

2 Theory

In the two-stub tuner, the separation between the stubs is fixed and the reactance of each stub is variable, while a slide-screw tuner has as variables the depth of the screw and the length of guide between screw and the impedance to be matched. For the double-slug tuner, the two degrees of freedom (Fig. 1) are the lengths of transmission line l and l' between the obstacles of fixed impedance A and B.

The reflection coefficient obtained from the tuner depends on the obstacles A and B and on the length of line l between them. It will be shown that the locus of the overall impedance Z^T obtained by varying the distance l between the obstacles describes a circle. The maximum value of reflection coefficient that may be generated depends of course on the impedances of the obstacles A and B and thus on their physical dimensions.

Having obtained the desired reflection coefficient, the phase angle necessary to the desired match depends on the length l' between the obstacle B and the load.

2.1 Derivation of tuner impedance Z^T

Consider the obstacle A in Fig. 1: it may be represented as a lossless, reciprocal network describable by a two-port scattering matrix $[S]^A$. Similarly, obstacle B may be represented by the two-port matrix $[S]^B$. The two obstacles are mounted in a homogeneous section of line of characteristic impedance Z_0, terminated in a nonreflecting load Z_0. The matrix of the section of line l plus obstacle B is given by

$$[S]^B = \begin{bmatrix} S_{11}^B\, e^{-2\gamma l} & S_{12}^B\, e^{-\gamma l} \\ S_{12}^B\, e^{-\gamma l} & S_{22}^B \end{bmatrix}$$

where the propagation constant $\gamma = \alpha + j\beta$, and α and β are, respectively, the attenuation and phase constants of the homogeneous transmission line.

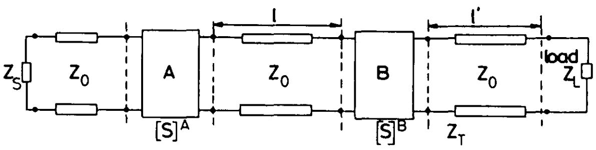

Fig. 1 *Equivalent circuit of double-slug tuner*

The overall output impedance Z^T may be obtained from the output reflection coefficient S_{22}^T of the overall network: obstacle A, line l and obstacle B.

$$S_{22}^T = \frac{S_{22}^B - \Delta^B\, S_{22}^A\, e^{-2\gamma l}}{1 - S_{11}^B\, S_{22}^A\, e^{-2\gamma l}} \tag{1}$$

where $\Delta^B = S_{11}^B\, S_{22}^B - S_{12}^B\, S_{21}^B$.

Consider the case where the transmission line as well as the obstacles are lossless (attenuation constant $\alpha = 0$). Eqn. 1 becomes

$$S_{22}^T = \frac{A - B\, e^{-2j\beta l}}{1 - C\, e^{-2j\beta l}} \tag{2}$$

where $A = S_{22}^B$, $B = \Delta^B\, S_{22}^A$ and $C = S_{11}^B\, S_{22}^A$. Since the parameters A, B and C are constant and, in general, complex quantities, expr. 2 is a bilinear equation. Now the locus of $e^{-2j\beta l}$ describes a circle of unit radius as l varies between 0 and $\lambda/2$. By the property of the bilinear transformation this circle will transform as a circle on the S_{22}^T plane, with centre c and radius r given by

$$c = \frac{-BC^* + A}{1 - |C|^2} \tag{3}$$

$$r = \frac{|AC - B|}{1 - |C|^2} \tag{4}$$

The locus of S_{22}^T is thus a circle with maximum value $|c| + |r|$ and minimum of $|c| - |r|$, for $|r|$ always less than or equal to $|c|$.

2.2 Transmission-line model

The most commonly used type of double-slug tuner is one in which low-impedance distributed line sections are used as the obstacles A and B. If one considers the case where discontinuities do not greatly perturb the field distribution, a TEM mode analysis may be used, and the scattering parameters of obstacle A obtained by treating it as a section of line of

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characteristic impedance Z_1 and length l_1 (Fig. 2).

$$S_{11}^A = S_{22}^A = \frac{\Gamma(1-e^{-2j\beta_1 l_1})}{1-|\Gamma|^2 e^{-2j\beta_1 l_1}} \tag{5}$$

$$S_{12}^A = S_{21}^A = \frac{(1-|\Gamma|^2)e^{-j\beta_1 l_1}}{1-|\Gamma|^2 e^{-2j\beta_1 l_1}} \tag{6}$$

where

$$\Gamma = \frac{Z_1 - Z_0}{Z_1 + Z_0}$$

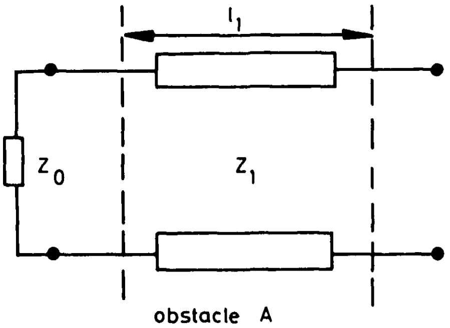

Fig. 2 *Model used for distributed slug analysis*

$$[S]^A = \begin{bmatrix} S_{11}^A & S_{12}^A \\ S_{21}^A & S_{22}^A \end{bmatrix}$$

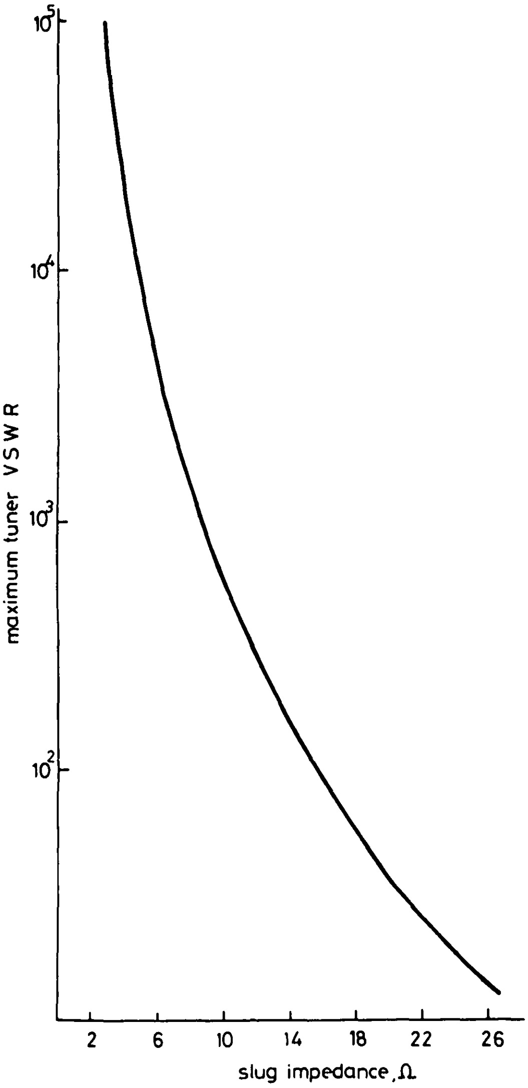

Fig. 3 *Maximum VSWR obtainable from tuner as a function of slug impedance (quarter-wavelength slugs)*

and β_1 is the phase constant of the low-impedance section A. Now if obstacles A and B are considered to be *identical*, the output reflection of the tuner, from eqn. 1, becomes

$$S_{22}^T = \frac{S_{22}^A(1-\Delta^A e^{-2j\beta l})}{1-(S_{22}^A)^2 e^{-2j\beta l}}$$

Thus, as l is varied from 0 to $\lambda/2$, the locus of S_{22}^T is a circle, with centre c and radius r given by

$$c = \frac{S_{22}^A - \Delta^A (S_{11}^A)^* |S_{22}^A|^2}{1-|S_{22}^A|^4} \tag{7}$$

$$|r| = |c| \tag{8}$$

From these equations the maximum tuner VSWR as a function of slug characteristic impedance Z_1 is obtained for impedance values of 3 Ω to 24 Ω, $l_1 = \lambda g/4$, (Fig. 3). However the TEM model is only valid when the discontinuities at the slug/transmission-line interface are negligible. Thus in practice tuners are designed with slug characteristic impedance greater than 10 Ω. From Fig. 4, one can obtain the bandwidth for a given

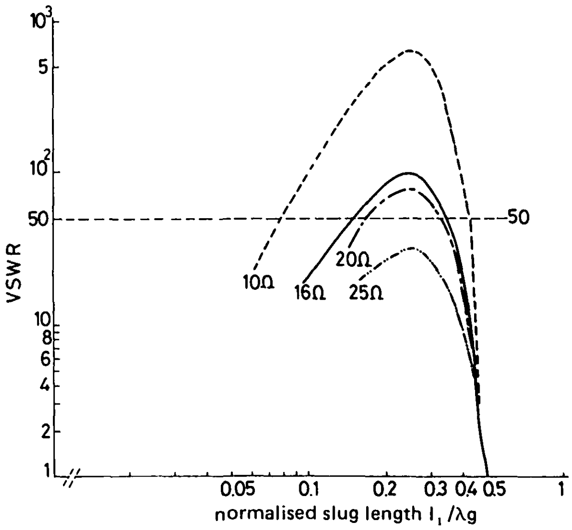

Fig. 4 *VSWR bandwidth obtainable for different values of slug impedance*

VSWR requirement for different values of slug characteristic impedance between 10 and 25 Ω. For example, for a specified VSWR of 50:1, a slug impedance of 10 Ω allows tuning over two and a half octaves (normalised slug length $l_1/\lambda g$ varying from 0.08–0.43), while for a transmission-line tuner using 16 Ω slugs (currently used in our laboratory for 4–8 GHz and 6–12 GHz tuners), octave band tuning is obtained.

2.3 Lumped-element model

The lumped-element model may be applied to the case of microstrip disc tuners, which are increasingly used in the characterisation of GaAs MESFET transistors. Provided that the dimensions of the discs used are less than an eighth of a wavelength at the frequency considered, they may be characterised as a lumped susceptance jB.

The centre of the locus of S_{22}^T and the radius are for this case given by

$$c = \frac{-b(b+2j)}{2(|b|^2+2)} \tag{9}$$

$$r = \frac{b\sqrt{|b|^2+4}}{2(|b|^2+2)} \tag{10}$$

where b is the normalised susceptance of the lumped capacitor, $b = jBZ_0 = j\Omega CZ_0$ and Z_0 is the line characteristic impedance.

2.4 *Effect of losses on match*

An important limitation of the performance of tuners, and one that is often neglected, is the effect of losses. As already seen, the impedances of the obstacles A and B and the length l of line between them determine the reflection coefficient generated by the 'intrinsic' tuner. The phase of the reflection coefficient is given by the line length l' between obstacle B and the load Z_L. In a similar manner it is convenient to separate losses into those generated within the 'intrinsic' tuner (Fig. 5*a*) and those generated in the line between tuner and load (Fig. 5*b*).

The slugs A and B in the loss analysis are treated as segments of transmission lines, as are the lengths of line l and l'. One may now apply the low-loss system model [7] and replace the imaginary propagation constant $j\beta_1$ in eqns. 5 and 6, by the complex propagation constant $\gamma_1 = \alpha_1 + j\beta_1$.

As the slugs used in the K-band tuner described in Section 3 are composite structures, it is difficult to evaluate the effect of the thermoretractable rubber sleeving on the phase constant β_1 of the overall slug transmission-line impedance. The solution

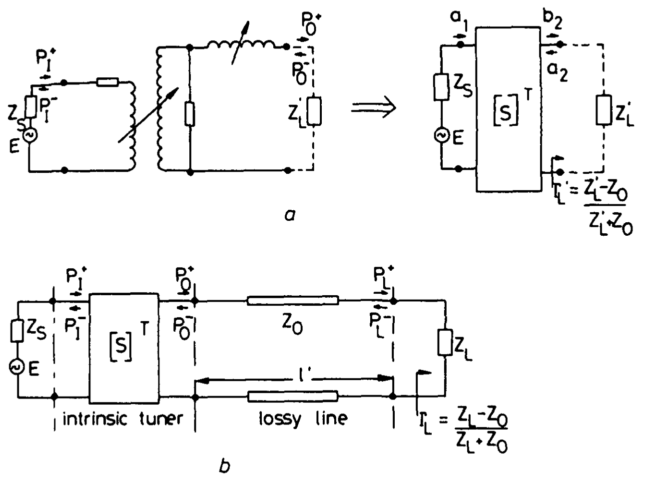

Fig. 5 *Tuner loss model*
a Loss model for intrinsic tuner
b Overall tuner loss model

adopted was to treat the slug as though it was composed of two line sections in series; one of characteristic impedance Z' between slug outer and the shielding, the other of impedance Z'' between slug inner and the centre conductor (Fig. 9). The calculation of the complex propagation constant γ for the line sections l and l' is straightforward.

Consider now the matrix $[S]^T$ with loss included, obtained by substituting the appropriate relationship for the slugs and transmission line l in eqns. 1. The efficiency of the 'intrinsic' tuner is defined as the ratio of the overall power P_o leaving the tuner to the overall power P_I entering it (Fig. 5*a*).

$$P_I = |a_1|^2$$
$$P_o = |b_2|^2 (1 - |\Gamma_L'|^2) \tag{11}$$

Thus, using the matrix $[S]^T$ to obtain the relationship between $|b_2|$ and $|a_1|$,

$$\eta = \frac{P_o}{P_I} = \frac{|S_{21}^T|^2 (1 - |\Gamma_L'|^2)}{[1 - |S_{22}^T|^2]^2} \tag{12}$$

where Γ_L' is the reflection coefficient of the effective load Z_L' seen by the intrinsic tuner. Define the loss factor A as the ratio of the power P_I^+ leaving the transmission line l' to that P_o^+ incident on it. Then

$$|\Gamma_L'|^2 = A^2 |\Gamma_L|^2 \tag{13}$$

and the overall power loss in the transmission line l' is (Fig. 5*b*):

$$L = \frac{P_L}{P_o} = \frac{P_L^+ - P_L^-}{P_o^+ - P_o^-} = \frac{(1 - |\Gamma_L|^2)A}{1 - A^2|\Gamma_L|^2} \tag{14}$$

The total loss in a tuner assuming perfect match between a source impedance Z_S and a load impedance Z_L is thus

$$\text{Loss} = P_2 P_I = \eta L = \frac{|S_{21}^T|^2 (1 - |\Gamma_L|^2)A}{[1 - |S_{22}^T|^2]^2} \tag{15}$$

Since an important part of the overall loss is made up by the conductor loss in the transmission lines, the overall loss (Fig. 6) is a function of frequency as well as reflection coefficient.

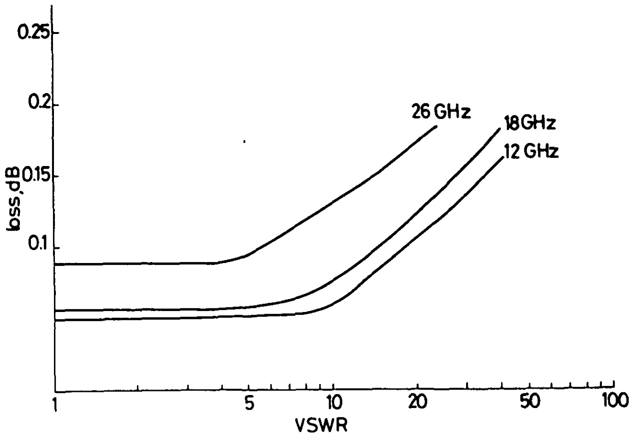

Fig. 6 *Tuner insertion loss as a function of VSWR calculated at several frequencies*

An important effect is to limit the maximum VSWR that can effectively be supplied by the tuner. This is shown by the curve of Fig. 7, in which VSWR taking into account loss is plotted against tuner VSWR which may be obtained from a lossless model at 12 GHz. It is seen that it is difficult in practice to obtain a VSWR greater than 40:1 above 12 GHz, which corresponds to a load reflection coefficient $|\Gamma_L| = 0.95$.

3 Tuner design

The tuner was specified around a need to obtain a maximum reflection coefficient of 0.9 across more than one octave band-width, from 12 to 26 GHz. This was to permit general-purpose utilisation of the tuners in experimental subassemblies

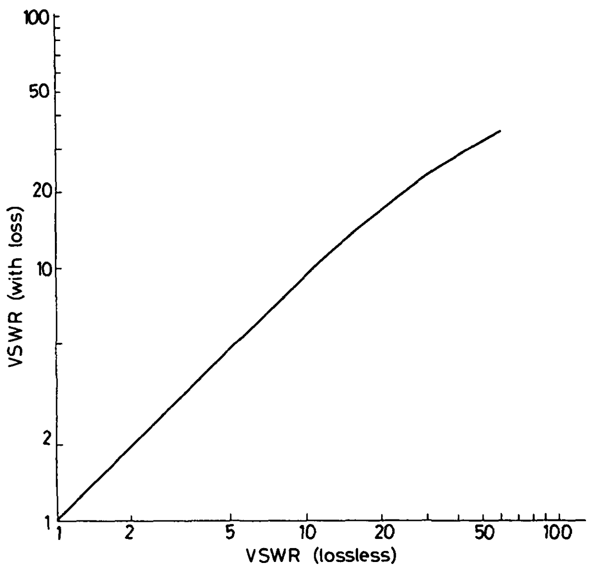

Fig. 7 *Effect of loss on VSWR obtainable from a K-band tuner at 12 GHz*

designed for the telecommunications bands 17.9–19.7 GHz and 21.2–23.5 GHz. In order to eliminate radiation losses it was decided to completely enclose the slugs by the transmission-line outer, and to control their movement by a magnet. To achieve this, it was necessary to use a thin-walled external conductor permeable to a magnetic field, soft iron slugs and miniature permanent magnets with a toroidal rubber drive mechanism (Fig. 8). To maintain mechanical strength, while

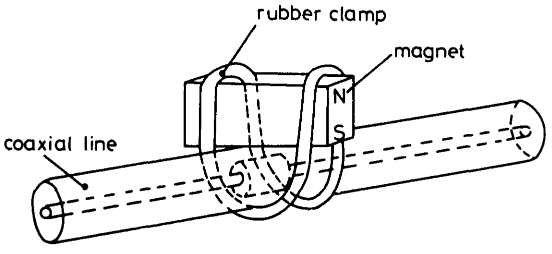

Fig. 8 *Schematic diagram of K-band tuner*

avoiding overmoding, a 4 mm internal-diameter stainless-steel tubing was selected for the outer conductor. To reduce conductor losses, the interior of the tubing was sanded down and plated up with 2 μm of gold.

To achieve the specification, it was observed from Fig. 4 that a lossless VSWR of 50:1 could be obtained across the desired frequency band by selecting a slug impedance of 16 to 18 Ω. From Fig. 7, it was noted that a VSWR of 50:1 (lossless) corresponded to a net VSWR of about 30:1 ($|\Gamma_L| = 0.94$) at 12 GHz. As the effect of loss on overall VSWR does not greatly change with frequency for these values of reflection coefficient, this value of slug impedance was retained. The slug length is calculated from Fig. 4 ($l/\lambda g = 0.15$ at 12 GHz) was 4 mm.

From a practical aspect, it was decided to guide the slug along the inner wall of the outer conductor. This was achieved by sheathing the outside of the slug in a thermoretractable plastic tubing which was then cured until the slug fitted snugly in the stainless-steel casing (Fig. 9). The dimensions of the slug outer and casing inner diameters were 3.4 mm and 4 mm, respectively, yielding a slug impedance of 6.7 Ω, the dielectric constant of the plastic sheath being $\epsilon_r = 2.1$, with a loss tangent of 0.002. The slug was then bored to an appropriate inner diameter hole (2.1 mm for a composite slug impedance of 18 Ω), to enable it to slide about the 1.74 mm diameter brass centre conductor. The slug impedance could thus be adjusted by varying the diameter of the bore. The 2.1 mm diameter bore was selected from practical considerations, an overall slug impedance of 16 Ω corresponding to a bore diameter of 2.03 mm.

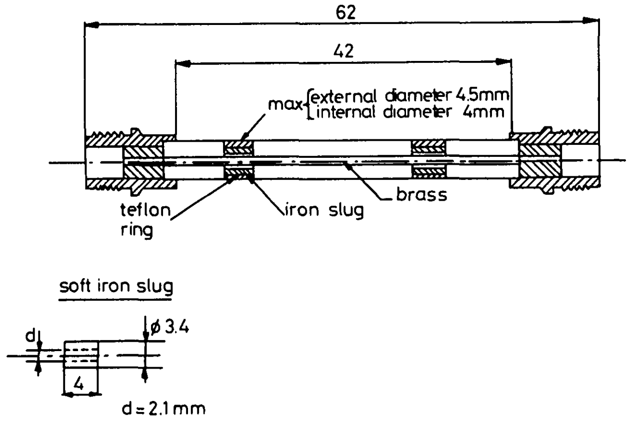

Fig. 9 *K-band tuner dimensions*

4 Experimental results

Prototype tuners were developed with three slug bore diameters: 2.1 mm, 2.3 mm and 2.4 mm. These correspond to overall slug impedances of 18 Ω, 23 Ω and 26 Ω. The measured values of maximum reflection coefficient at 12, 18 and 24 GHz are given in Fig. 10 for the 26 Ω tuner, while all three tuners are measured at 12 GHz. It is seen that a reasonably good agreement is obtained between the measured results and those obtained from the model including loss.

An important source of error, particularly at the higher frequencies, was the neglect of the mismatch between the coaxial line and the 3 mm connector, as well as neglect of all connector-introduced losses. It was also found in practice that the overall slug length should have been reduced by about 6% in order to take into account the effect of the plastic sleeving. A new tuner version with APC 3.5 connectors is currently being developed.

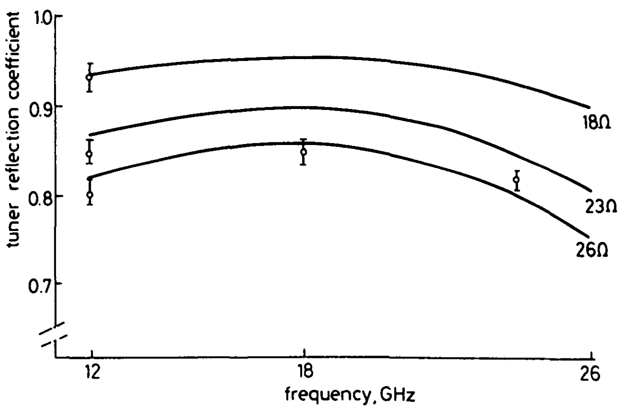

Fig. 10 *Performance of K-band tuner for different slug characteristic impedances*

——— calculated

measured

5 Conclusion

An analysis of double-slug tuners was developed and applied to two cases: a distributed transmission-line model and a lumped-capacitance model. The transmission-line model was developed to include the effect of losses both of the intrinsic tuner and in the length of line connecting the tuner to the external load. The curves were used in the design of a *K*-band tuner of the frequency range 12–26 GHz. The tuner was realised in a 4 mm internal diameter stainless steel tubing with magnetically guided slugs. Good agreement was obtained between calculated and measured results.

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References

1 RAGAN, G.L.: 'Microwave transmission circuits' (McGraw-Hill, 1948) vol. 9, pp. 406–472
2 PAUCKSH, H., SCHNEIDER, H., and SCHNEIDERS, H.: 'Aufbau and Eichung eines Höchstfrequenztransformators', *Fern. Zeit.*, 1954, (6), pp. 281–284

3 LUXTON, H.E.G.: 'Gallium arsenide field effect transistors – their performance and application up to X-band frequencies', Proceedings of fourth European microwave conference, Montreux 1974, pp. 92–96
4 SOARES, R.A., and CRIPPS, S.C.: 'Some novel characterization techniques used in the design of X-band GaAs FET amplifiers'. Proceedings of sixth European microwave conference, Rome 1976, pp. 19–23
5 SOARES, R.A.: 'Novel large signal S-parameter measurement technique aids GaAs power amplifier design'. Proceedings of seventh European microwave conference, Copenhagen 1977, pp. 113–117
6 de Los REYES DAVO, E.: 'Estudios de amplificadores de microondas con dispositivos GaAs MESFET'. Doctoral thesis, Escuela Tecnica Superior de Ingenieros de Telecommunicacion, Universidade Politecnica de Barcelona, Sept. 1979
7 COLLIN, R.E.: 'Foundations for microwave engineering' (McGraw-Hill Kogakusha, 1966) pp. 76–80
8 LE BAIL, J.L., and DA ROCHA,: 'Mesure des transistors hyperfréquences de puissance et réalisation d'un amplificateur 2 W – 2 GHz'. Note technique EST/EFT 511, CNET, Paris 1973